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Electrical post-compensation of intrachannel nonlinearities in 10GBaud coherent QPSK transmission systems

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ABSTRACT

Electrical post-compensation of intrachannel nonlinearities in 10GBaud coherent QPSK transmission systems is proposed. The channel inversion method is implemented with a coarse split-step Fourier algorithm through digital signal processing at the receiver side. Simulation results show that simultaneous compensation of intrachannel nonlinearities and chromatic dispersion can be achieved with much simplified signal processing structure. The required optical signal-to-noise ratio at certain bit error ratio is remarkably reduced when the intrachannel nonlinearities induced impairments are dominant.

1. Introduction

In recent years, with the development of high-speed digital signal processing (DSP) and analog-to-digital conversion (ADC), coherent detection is revisited and attracts extensive studies [1]. DSP technology simplifies coherent detection and brings it more advantages such as flexibility, hardware transparence and potential to adaptively compensate for channel impairments in electrical domain [2–4]. It has been reported that polarization-multiplexed QPSK signal of 42.8 Gbit/s (10.7GBaud) is transmitted over 6400 km standard single mode fiber (SSMF) and the digital coherent receiver allows chromatic dispersion (CD) compensated totally in electrical domain without using inline optical dispersion compensator (ODC) [3]. The lumped dispersion compensation scheme makes the fiber link simple, flexible and potentially cost effective. Such dispersion mapping can also alleviate inter-channel nonlinearities [5], which is quite difficult to eliminate digitally. Therefore the fiber link with full electrical dispersion compensation is more appropriate for long haul coherent detection systems.

However, such pseudo-linear long haul transmission systems are greatly degraded by intrachannel nonlinearities even at 10GBaud due to high accumulated inline dispersion [6]. In direct detection systems electrical pre-distortion (EPD) can be used to equalize intrachannel nonlinearities which is realized by modeling the channel inversion (CI) for nonlinear effects and pre-distorting the transmitted waveform accordingly [7,8]. A fast algorithm for electrical pre-equalization of self-phase modulation and chromatic dispersion in 10 Gbit/s RZ-OOK systems has been presented in [9]. With both amplitude and phase information available at the receiver, for coherent detection, it is more convenient to eliminate transmission impairments by post-compensation using available DSP technique [10–13]. Ref. [13] shows that multi-channel joint back-propagation can post-compensate both inter-channel and intrachannel impairments in 10×10 Gbit/s BPSK wavelength division multiplexing (WDM) systems. However, it is difficult to implement in practice because it requires the data from other channels. Moreover, fiber links without inline ODCs are generally used in coherent systems as has been mentioned. In the following we'll prove that for such fiber links intrachannel four wave mixing (IFWM) other than cross phase modulation (XPM) plays the dominant role for nonlinear degradations in WDM systems.

Aiming at mitigating intrachannel nonlinearities, in this paper we implement a simplified CI algorithm in 10GBaud coherent NRZ-QPSK systems with cascaded SSMF link. The paper is organized as follows. In Section 2, we present the system configuration and the CI method in detail. Section 3 demonstrates that for single-channel operation over such fiber link without inline ODCs, IFWM is the dominant impairment while nonlinear phase noise (NPN) can be safely neglected. In Section 4, simulation results show both intrachannel nonlinear impairments and chromatic dispersion can be compensated efficiently and the computation complexity is greatly reduced with the simplified CI method. In Section 5, we explore the impact of both laser linewidth and the frequency offset between the transmitter and the local oscillator (LO). The

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application in WDM systems is also discussed. Finally we draw our conclusions in Section 6.

2. System configuration and CI method

We consider a single-channel 10GBaud NRZ-QPSK coherent communication system as shown in Fig. 1. A pair of lithium niobate ($LiNbO_3$) dual-parallel Mach–Zehnder modulators is used for QPSK modulation. The transmitted signal is assumed with constant polarization, with which the LO light is aligned. The fiber link consists of N amplified SSMF spans for transmission, without inline ODCs. The gain of erbium-doped fiber amplifier (EDFA) equals the span loss. DSP based phase diversity receiver is employed. After 2×4 optical hybrid and balanced detection, ADC is performed for I and Q branch, respectively. Then electrical CI equalization is applied to eliminate transmission impairments. Finally the equalized signals are sent into Viterbi-and-Viterbi carrier phase estimation module to retrieve the encoded phase [2].

CI equalization is modeled by numerically solving the nonlinear Schrödinger equation with negative link parameters using symmetric split-step Fourier method [7,8]. In each step, the optical field $A(z,t)$ evolves according to Eq. (1) [14]

$$\begin{aligned} &A(z+\Delta h, t) \\ &\approx \exp\left(\frac{\Delta h}{2}\widehat{D}^{-1}\right)\exp\left(\int_z^{z+\Delta h}\widehat{N}^{-1}(z')dz'\right)\exp\left(\frac{\Delta h}{2}\widehat{D}^{-1}\right)A(z,t) \\ &\approx \exp\left(\frac{\Delta h}{2}\widehat{D}^{-1}\right)\exp(\Delta h\widehat{N}^{-1})\exp\left(\left(\frac{\Delta h}{2}\widehat{D}^{-1}\right)\right)A(z,t) \qquad (1) \end{aligned}$$

$\widehat{D}^{-1}$ and $\widehat{N}^{-1}$ are differential and inverse nonlinear operators respectively. Δh is the step length.

However, the virtual back-propagation usually needs extensive computation which is time-consuming. We thus apply a coarse split-step Fourier algorithm in CI processing to improve computation efficiency. In our simulation the step length is adaptively determined by the approximated maximum nonlinear phase shift $\Delta\Phi_{max} = \gamma P \Delta h(180/\pi)$. γ is the fiber nonlinear coefficient. P is the optical power along the fiber link and Δh is alterable controlled by $\Delta\Phi_{max}$. The computation can be sped up by reducing step number, i.e., adopting larger step length. When launch power is fixed, large Δh leads to increased $\Delta\Phi_{max}$ and fast computation. Meanwhile, the compensation is less accurate and higher optical signal-to-noise ratio (OSNR) is needed to get the same performance compared with that for small Δh. In the following we will demonstrate that the fast algorithm can be implemented with acceptable OSNR margins.

3. Intrachannel nonlinearities and nonlinear phase noise

The intrachannel nonlinear impairments in phase-modulated systems operating in the pseudo-linear regime mainly include IFWM and NPN caused by the interaction of amplified spontaneous emission (ASE) and fiber nonlinearities. For 40 Gbit/s RZ-DPSK transmission, it has been demonstrated that IFWM is the dominant impairment in fiber link with inline residual dispersion per span [15]. In 10GBaud transmission system, similar results can be concluded because the residual dispersion in each span can also introduce de-correlation to NPN and thus suppress it, while IFWM accumulates span by span. To prove this point, we carry out numerical simulations using commercial software VPItransmissionMaker 7.5.

We choose a De Bruijn sequence of 2^{14} bits to capture the nonlinear interaction details in our system scenario. The parameters of SSMF are as follows: span length $L = 80$ km, attenuation $\alpha = 0.21$ dB/km, dispersion $D = 17$ ps/nm/km and nonlinearity $\gamma = 1.32$ W/km; span number $N = 100$. The mean nonlinear phase shift $\langle\Phi_{NL}\rangle = N\gamma P_{in} L_{eff}$ is used to represent the nonlinear strength. P_{in} is the average launch power and $L_{eff} \approx 1/\alpha$ is the effective nonlinear length per span.

The inline EDFA is set with and without ASE noise, respectively. Launch power P_{in} is set to -4.27 dBm (corresponding to $\langle\Phi_{NL}\rangle = 1.0$ rad). The OSNR of the received signal is 12.93 dB with 4.0 dB EDFA noise figure (NF) assumed. For the case of noise free EDFA, ASE noise is added before the receiver to keep the same OSNR value. The linewidth of the transmitter and the LO are both 1.0 MHz, which is a typical value for semiconductor distributed feedback (DFB) lasers. In Fig. 2a and b, we compare the received constellation diagrams after carrier phase estimation and the bit-error-rate (BER) for the cases of without and with NPN, respectively. As expected, the phase noise is quite close to each other, demonstrating the fact that NPN is quite small compared to IFWM. The little difference between the BER values proved that we can reasonably neglect NPN for convenience without degrading the validity of our conclusion.

4. Simulation results

In this section we numerically demonstrate the efficiency of the simplified CI processing for mitigating intrachannel nonlinearities. According to the conclusion of previous section, NPN is not considered. The inline amplifier is set noiseless and ASE noise is added before the coherent detector to set different OSNRs. The system performance is specified in terms of the required OSNR to achieve a BER target of 10^{-3} for a resolution bandwidth of 0.1 nm. BER is calculated using Monte Carlo method with more than 100 errors counted.

In DSP based coherent receiver, digital equalization usually requires two samplings per symbol and the analog signal has to be digitized [2–4]. To study the quantization impact on CI processing, we calculate the OSNR penalty for different numbers of quantization bit referencing to the ideal continuous case. The results are shown in Fig. 3. Note that for the CI algorithm in Fig. 3, $\Delta\Phi_{max}$ is set as 0.05°, equaling to the value for modeling practical transmission. It can be seen that using six-bit quantization yields a negligible OSNR penalty. Therefore, we choose six-bit quantization together with two samplings per symbol in our simulations.

We compare two equalization methods-finite impulse response (FIR) filtering and CI processing in Fig. 4a. The tap number and coefficients of FIR filter are determined according to the Fourier transform of the inverse fiber dispersion function [2–4]. A FIR filter

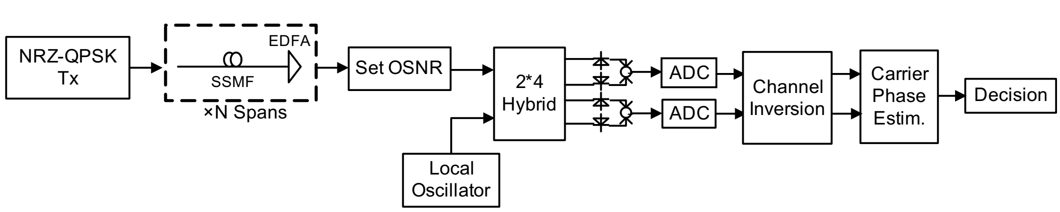

Fig. 1. Schematic of coherent NRZ-QPSK transmission system with CI processing.